On the End-to-End Distortion for a Buffered Transmission over Fading Channel

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Abstract

In this paper, we study the end-to-end distortion/delay tradeoff for a analogue source transmitted over a fading channel. The analogue source is quantized and stored in a buffer until it is transmitted. There are two extreme cases as far as buffer delay is concerned: no delay and infinite delay. We observe that there is a significant power gain by introducing a buffer delay. Our goal is to investigate the situation between these two extremes. Using recently proposed *effective capacity* concept, we derive a closed-form formula for this tradeoff. For SISO case, an asymptotically tight upper bound for our distortion-delay curve is derived, which approaches to the infinite delay lower bound as $\mathcal{D}_\infty \exp(\frac{C}{\tau_n})$, with τ_n is the normalized delay, C is a constant. For more general MIMO channel, we computed the distortion SNR exponent – the exponential decay rate of the expected distortion in the high SNR regime. Numerical results demonstrate that introduction of a small amount delay can save significant transmission power.

I. INTRODUCTION

Quality-of-Service (QoS) is a critical design objective for next-generation wireless communication system. In general, the data, voice and multimedia transmission over packet cellular networks, wireless LAN or sensor networks involves the analogue observations are transmitted to the end user over a wireless link. End-to-End distortion and transmission delay are two fundamental QoS metrics. Such QoS requirements pose a challenge for the system design due to the unreliability and time varying nature of the wireless link.
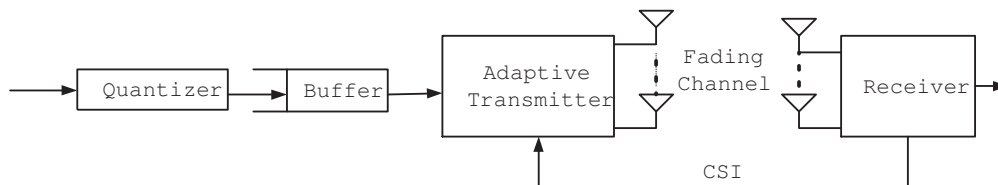

Fig. 1. System model

In this paper, we consider transmission of an analogue source over a wireless time-varying fading channel. Our goal is to optimize the end-to-end distortion given a delay constraint. We first focus on the single antenna case (SISO) and derive the distortion and delay tradeoff for the wireless fading channel. We then extended our model to multiple input and multiple output (MIMO) block Rayleigh fading channel. We compute the SNR exponent [1] for the buffered transmission. To this end, we adopt a cross-layer approach shown in Figure 1. At this point, for simplicity we assume an independent and identically distributed (i.i.d.) block fading channel model. Such a model is suitable for serval practical communication scenarios, e.g., time hopping in TDMA, frequency hopping in FDMA and multicarrier systems. Extension to more practical time-correlated case will be discussed later. Throughout this paper, we always assume channel state information (CSI) is perfectly known at the receiver and the transmitter only know the instantaneous channel capacity via a feedback link (transmitter don't need to know the exact channel realization).

We consider an i.i.d. complex memoryless Gaussian source $\sim \mathcal{CN}(0, 1)$, which is quantized it and then fed into a buffer. Since the channel is time-varying, the transmitter adjust the transmission rate to the current channel status. The relevant performance criteria are the end-

to-end quadratic distortion and the buffer delay. We aim to find the relationship between the distortion and delay for some average transmission power. The Gaussian source is a good approximation of more general source distribution in high resolution regime [3], [4]. We assume that each group of K source samples is tranmsitted over N channel uses on average. We define the corresponding *bandwidth ratio* as

$$\eta = \frac{N}{K},\tag{1}$$

where K is large enough to consider the source as ergodic and N is large enough to design codes that can achieve the instantaneous channel capacity. Our tools here are the large deviation theory and information theory.

Recently, some researchers have considered such end-to-end quadratic distortion as the performance criteria. In [2], Holliday and Goldsmith first investigated the end-to-end distortion for the MIMO block fading channel, based on the source-channel separation theorem and Zheng and Tse's diversity-multiplexing trade-off. And they also incorporated the delay consideration into their model using ARQ argument, which is different from our approach. In [3], Laneman et al., considered the problem of minimum average distortion transmission over parallelled channels. They introduced the distortion SNR exponent as a figure of merit for high SNR value, and compared the multiple description source coding diversity and channel coding diversity. Caire and Narayanan [1] pointed out the the separation theorem does not hold for delay constrained and the unknown channel at the transmitter end, they investigated the SNR exponent of the distortion function in high SNR regime for this problem, an upper bound and lower-bound for the distortion SNR exponent were derived. [4] Gunduz and Erkip extended their results by a layered broadcast transmission scheme. For some bandwidth ratio, the optimum SNR exponent is achieved.

For the combination of queuing and information theories, in [5], Wu and Negi, first proposed the concept of effective capacity, which is an extension of Shannon's capacity by incorporating into the buffer delay. The effective capacity is the dual of the Chang's effective bandwidth [7] in the network literature. Negi and Goel [6] unified the effective capacity with error exponent for more practical considerations. A QOS-aware rate and power control algorithm for wireless fading channel was proposed by Tang and Zhang [8].

For buffered transmission, Berry and Gallager investigated the power and delay tradeoff for

communication over fading channel [9]. In [10], Tse analyzed the distortion for a fixed line networks, but with adaptive quantizer.

The rest of this paper is organized as follows: in Section II, we state the problem and show inserting a buffer can save significant power. We introduce the system model and some preliminaries of the effective capacity in Section III. Section IV develops our main results–distortion-delay function and an upper bound for SISO channel, some asymptotic analysis is provided. In Section V, We extend the distortion analysis to MIMO channel, and the SNR exponent for buffered transmission is derived. Distortion-delay for large antenna MIMO channel is also derived by utilizing the mutual information Gaussian approximation. Finally, Section VII concludes the paper.

Throughout this paper, normal letters indicate scalar quantities and boldface fonts denote matrices and vectors. For any matrix $\mathbf{M}$ we write its transpose as $\mathbf{M}^T$ and $\mathbf{M}^H$ is its conjugate transpose. x^* denotes the conjugate of x. $\ln(\cdot)$ and $\log(\cdot)$ represent the natural and 2 based logarithm.

II. PROBLEM STATEMENT

For buffered transmission over the fading channel, there are two extreme cases: 1) There is no buffer — no delay, 2) we have an infinite buffer size, i.e., we allow an infinite transmission delay. For the first case, we adaptively quantize the Gaussian source according to the CSI. Assuming perfect transmission, we can approximate the average achievable quadratic distortion by:

$$\mathcal{D}_0(\rho) = \mathrm{E}[\exp(-\eta \ln(1 + |h|^2 \frac{P}{N_0 W}))] \; , \tag{2}$$

where P denotes the transmission power, W and N_0 resent the bandwidth and noise variance; h is the channel gain, a random variable with unit variance follow a certain statistical distribution. Here, we have used the information theoretical results: Gaussian distortion-rate function can be express as $\mathcal{D}(R_s) = \exp(-\eta R_c)$ and $C(\rho) = \log(1 + |h|^2 \rho)$ is the instantaneous channel capacity-cost function. For infinite delay case, the average transmission rate can achieve the ergodic capacity of a fading channel and the quantizer can simply adopt a constant output rate. The average distortion is given by:

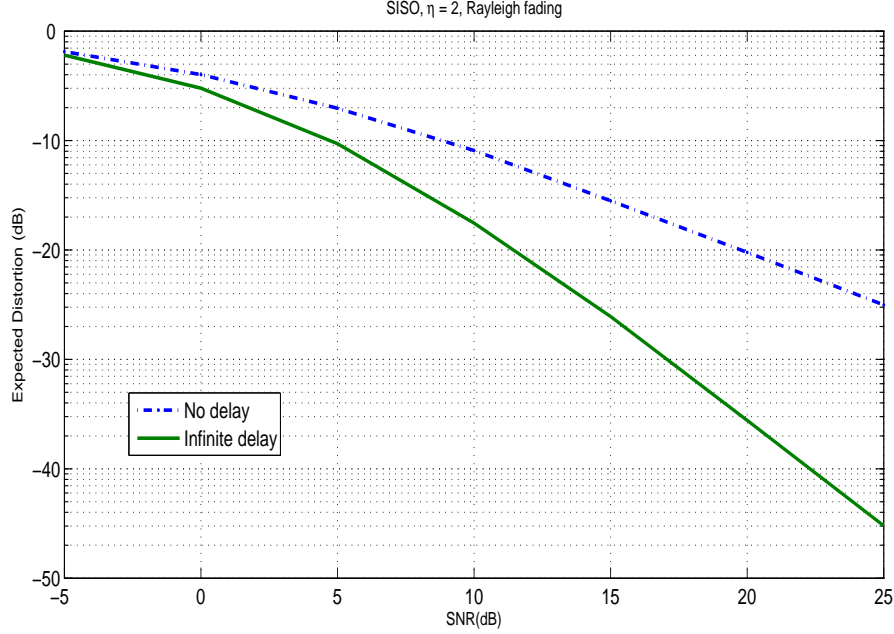$$\mathcal{D}_\infty(\rho) = \exp(-\eta \, \mathrm{E}[\ln(1 + |h|^2 \frac{P}{N_0 W})]) \; . \tag{3}$$

Fig. 2. Distortion of Gaussian Source Transmitted over i.i.d. Rayleigh fading Channel.

The function $\exp(-(\cdot))$ is a covex function. Due to Jensen's inequality, the distortion $\mathcal{D}_0$ is low bounded by $\mathcal{D}_\infty$, i.e., $\mathcal{D}_0 \geq \mathcal{D}_\infty$. The two distortion functions are plotted in Figure 2 for a Rayleigh fading channel. Notice that there is a gap between no-delay and infinite delay curves. We can call this transmission power gap as "Jensen's gain". Note, we assume $\eta = 2$ and a complex Gaussian source, this is equivalent to a real source with bandwidth ratio of one. So introducing a buffer at the transmitter to match the source rate with the instantaneous quality of the channel can save lots of transmission power to meet some distortion requirement. Also, we have simplified the quantization step (constant rate). A natural question is therefore: if we only allow a finite delay or buffer, how much gain can we achieve? How fast does the distortion curve converge to the infinite-delay lower bound as the delay increases? One of the the main result of this paper is a clear characterization the tradeoff between end-to-end quadratic distortion and delay, which provides insights to the impact of the buffer delay on the achieved distortion function of the memoryless analogue source transmitted over a wireless fading channel.

To answer the question raised earlier, we combine the ideas from the fields of queuing theory and communication/information theory to analyze the above problem. The tool we use here is the

concept of *effective capacity* [5], which is the dual of *effective bandwidth* in networking literature. The effective capacity synthesizes the channel statistics and QoS metric (delay and buffer overflow) into a single function using large deviation theory. It is a powerful and unified approach to study the statistical QoS performance of wireless transmission where the service process is time-varying. For i.i.d. SISO block fading channels we derived a closed-form expression for the distortion-delay curve, which is hard to analyze due to some mathematically intractable special functions. Then we give out a tight upper bound for this distortion-delay function to theoretically and asymptotically analyze the convergence behavior.

In Fig 2., we find the power gain is marginal for low SNR. As the SNR value increases, the gain becomes significant. This is because the $\exp(\cdot)$ and $\log(\cdot)$ functions are approximately linear in the low SNR regime. Hence, the "Jensen's gain" is negligible at low SNR. We can view the slope of the distortion–SNR curve as a similarity of the diversity order for the bit error rate in the wireless communication. Therefore, we will investigate the distortion SNR exponent for a buffered transmission. Introducing a buffer can provides some kind of time diversity. For the MIMO channel, besides the time diversity, we also have space diversity. We will look into the interplay between these two diversities and the impact of buffer on the SNR exponent.

III. SYSTEM MODEL

The system model is illustrated in Figure 1. We have an i.i.d. complex Gaussian source $\sim \mathcal{CN}(0, 1)$ with total bandwidth B_w. We quantize the source samples using vector quantizer or trellis coded quantizer (TCQ). The quantization operate every K samples a time and fed into a buffer with size B bits. Let the K samples have time duration T_f, so each frame have $T_f \times B_w \times R_s = K \cdot R_s$ bits, where R_s bits is number of bits into which each Gaussian sample is quantized. K is large enough to ensure ergodic of the source.

We assume a MIMO i.i.d. block fading channel with M_t transmit and M_r receive antennas. The SISO, MISO and SIMO are special cases of this general model. The channel model can be expressed as:

$$\mathbf{y}_i = \sqrt{\frac{\rho}{M_t}} \mathbf{H} \mathbf{x}_i + \mathbf{w}_i, \qquad i = 1, \cdots, N \tag{4}$$

Where $\mathbf{H}$ is the channel matrix containing i.i.d. elements $h_{i,j} \sim \mathcal{CN}(0, 1)$ (Rayleigh independent fading). $\mathbf{x}_i$ is the transmitted signal at time i, the codeword $\mathbf{X} = [\mathbf{x}_1, \cdots, \mathbf{x}_N] \in \mathbb{C}^{M_t \times N}$ is

normalized so that is satisfies $\text{tr}(\mathbb{E}[\mathbf{X}^H\mathbf{X}]) \leq M_tN$. ρ denotes the signal-to-noise ratio (SNR), defined as the ratio of the average received signal energy per receiving antenna to the noise per-component variance. $\mathbf{Z} = [\mathbf{z}_1, \cdots, \mathbf{z}_N] \in \mathbb{C}^{M_r \times N}$ is the complex additive Gaussian noise with i.i.d. entries $\mathcal{CN}(0,1)$. We define $M_* = min(M_t, M_r)$ and $M^* = max(M_t, M_r)$.

A. Effective Capacity

The key idea of effective capacity is that, for a dynamic queuing system with stationary ergodic arrival and service process, the queue length $Q(t)$ converges in distribution to a random variable $Q(\infty)$. The probability of queue length exceeding a certain threshold B decays exponentially fast as the threshold B increases [5]. Mathematically,

$$\lim_{B \to \infty} \frac{-1}{B} \ln Pr\{Q(\infty) > B\} = \theta \ , \tag{5}$$

where θ is the QoS parameter decided by the delay requirement of the queue system. A large value of θ leads to a stringent delay requirement, i.e., small delay. In particular, as θ goes to ∞, the system can not tolerate any delay. On the other end, when θ goes to 0, the system can tolerate an arbitrarily delay.

Let the sequence $\{R[i], i = 1, 2, \ldots\}$ denote the discrete-time instantaneous channel capacity, which is a stationary and ergodic stochastic process. Define

$$S[t] \triangleq \sum_{i=1}^{t} R[i] \tag{6}$$

as the accumulate service provided by the channel. Assume the Gärtner-Ellis limit of $S[t]$:

$$\Lambda_C(\theta) \triangleq \lim_{t \to \infty} \frac{1}{t} \ln \mathrm{E}\left\{e^{\theta S[t]}\right\}, \quad \forall \theta > 0 \tag{7}$$

exits and is a convex function differentiable for all real θ. Then, the effective capacity with delay constraint decided by θ is defined as

$$E_C(\theta) \triangleq -\frac{\Lambda_C(-\theta)}{\theta} = -\lim_{t \to \infty} \frac{1}{\theta t} \ln \mathrm{E}\left\{e^{-\theta S[t]}\right\} \ . \tag{8}$$

In particular, for i.i.d. cases, the effective capacity simply reduces to the ratio of log-moment generating function of the instantaneous channel capacity to the exponent θ

$$E_C(\theta) = -\frac{1}{\theta} \ln \mathrm{E}\left\{e^{-\theta R[t]}\right\} \ . \tag{9}$$

The effective capacity falls into the large deviation framework, which is asymptotically valid for a large queue size.

IV. DISTORTION-DELAY FUNCTION

We will derive the closed-form expression for the end-to-end quadratic distortion given the delay constraint in this section. The starting point is vector quantization and delay bound violation probability using effective capacity. For a Gaussian source vector $\mathbf{u}$ with K samples that has support on $\mathbb{C}^K$, a KR_s-nats quantizer is applied to $\mathbf{u}$ via a mapping $\mathbf{u} \to \tilde{\mathbf{u}}$. The cardinality of discrete set $\tilde{\mathbf{u}}$ is e^{KR_s}. Define the average quadratic distortion by

$$\mathcal{D}_Q(R_s) \triangleq \frac{1}{K}\mathbf{E}[|\mathbf{u} - \tilde{\mathbf{u}}|^2] , \tag{10}$$

where the expectation is with respect to $\mathbf{u}$. According to the distortion-rate theory, the distortion function $D_Q(R_s) = \exp(-R_s)$ is achievable for a complex Guassian source. The quantized bits are transmitted over a statistical channel, let P_e denote the error probability of this channel. It has been shown in [12] that the achievable end-to-end distortion for such tandem scheme is upper bounded by

$$\mathcal{D}_{e-e}(R_s) \leq \mathcal{D}_Q(R_s) + O(1)P_e . \tag{11}$$

For our problem, if we assume using Gaussian code to achieve the instantaneous capacity, the delay bound violation (buffer overflow) probability will dominate the decoding error probability. From the effective capacity theory, we have the following approximation for P_e:

$$P_e \triangleq P_r\{Q(\infty) \geq B\} \approx \kappa e^{-\theta B} , \tag{12}$$

where θ is the QoS parameter, B is the buffer size; κ is a constant that denotes the probability that the buffer is non-empty. κ is large compared with P_e. Given the delay constraint at τ seconds, using Little's theorem, we have following result: $B = Rs \times B_w \times \tau$. B_w is the source bandwidth. Substitute (12), B and $D_Q(R_s)$ into (11), we may write the bound on the end-to-end distortion as

$$\mathcal{D}_{e-e}(R_s) \leq \exp(-R_s) + O(1)\kappa \exp(-\theta B_w R_s \tau) . \tag{13}$$

In order to get analytical results, we consider the asymptotically large delay and high SNR regime, i.e., small distortion. We can optimize the end-to-end distortion by choosing the two exponents equal to each other (exponential order tight). As a result, we have $\theta = \frac{1}{B_w \tau}$.

If we assume the transmitter don't know the channel realization, but know the value of instantaneous capacity via the feedback link. The instantaneous capacity can be achieved by the Gaussian codebook. We have following theorem.

Theorem 1: Given a delay $\tau = \frac{1}{B_w\theta}$ and bandwidth raio η, the distortion upper bound function of the i.i.d MIMO block fading channel can be expressed as:

$$\mathcal{D}(\theta) \leq \left[\mathbf{B}^{-1} \det[\mathbf{G}(\theta)] \right]^{\frac{1}{K\theta}} . \tag{14}$$

where $\mathbf{B} = \prod_{i=1}^{M_*} \Gamma(d+i)$, and $d = M^* - M_*$. And $\mathbf{G}$ is $M_* \times M_*$ Hankel matrix whose (i,j)th entry is defined to be

$$g_{i,j} = \int_0^\infty \left(1 + \frac{\rho}{M_t}\lambda \right)^{-\theta K\eta} \lambda^{i+j+d}e^{-\lambda}d\lambda, \quad i,j = 0,\cdots,M_* - 1 . \tag{15}$$

Γ is the complete Gamma function.

Proof: The Mutual information for the each MIMO block transmission can be expressed as:

$$R_s(\mathbf{H}) = K\eta \cdot \ln \det \left(\mathbf{I} + \frac{\rho}{M_t}\mathbf{H}\mathbf{H}^H \right) \tag{16}$$

plug into equation (9) and (13), we have

$$\mathcal{D}(\theta) \leq \left\{ \mathrm{E}\left[\det \left(\mathbf{I} + \frac{\rho}{M_t}\mathbf{H}\mathbf{H}^H \right) \right]^{-\theta K\eta} \right\}^{\frac{1}{\theta K}}$$

$$= \left\{ \int_0^\infty \prod \left(1 + \frac{\rho}{M_t}\lambda_i \right)^{-\theta K\eta} f(\boldsymbol{\lambda})d\boldsymbol{\lambda} \right\}^{\frac{1}{\theta K}} . \tag{17}$$

Where $0 \leq \lambda_1 \leq \cdots \leq \lambda_{M_*}$ denote the ordered eigenvalues of $\mathbf{H}\mathbf{H}^H$. The joint distribution of the λ_i's follows the Wishart pdf given by

$$f(\boldsymbol{\lambda}) = K_{M_t,M_r}^{-1} \prod_{i=1}^{M_*} \lambda_i^{M^*-M_*} \prod_{i<j}(\lambda_i - \lambda_j)^2 \exp\left(-\sum_i \lambda_i \right) , \tag{18}$$

where K_{M_t,M_r} is a normalization constant. Follow the results of [17], we can get the distortion function as (14). ∎

Remarks

- If we assume the quantization process is independent of the channel status, we can show the the constant quantization rate is the optimum one. First, for a buffered system with independent arrival and departure processes, the constant arrival processe is optimal with respect to the buffer overflow probability, for all the arrival processes that have the same average rate [7]. Second, given a buffer overflow probability, constant rate quantization will minimize the distortion according to the Jensen's inequality. Therefore, constant rate quantization is optimal if the quantization process is independent of the channel mutual

information. Another advantage of constant rate quantization is to reduce the quantizer design complexity.

- When the quantizer rate selection is according to the buffer state status. We can not prove the constant rate quantization is optimal. Hence the distortion of (14) is an upper bound. One extreme case is that the quantizer is chosen to make sure no buffer overflow, i. e. , the quantization rate selection is to match the channel mutual information profile. This scheme will degenerate to no buffer (delay) case. Therefore, it is serious suboptimal. The optimal quantizer rate should balance the "Jensen's gain" and the reduced distortion by decreasing the buffer overflow probability via quantization rate matching the buffer status.

The introducing buffer delay in (17) can be viewed as first shrinks the integrand near to 1 as $\theta \to 0$, and then restore it after taking the expectation. From Fig. 3, we can observe that after the contraction function of $(\cdot)^\theta$, as θ goes to zero, the integrand function become more linear. This observation can explain why we have a large gain after introducing a buffer delay mathematically, and provide some intuitions of distortion–delay function. Moreover, Fig. 3 shows that the large the bandwidth ration η, the more effective of the shrink operation (larger gain). Therefore, introduce a buffer delay has larger gain for high bandwidth ratio scenario, or high resolution quantization. We will confirm the result later theoretically by deriving the SNR exponent.

The result of Theorem 1. is very complicate, not so much insight can be given from the expression itself. In the ensuing part of this paper, we will first investigate the distortion-delay of SISO, MISO / SIMO case, which a simpler form can be arrived. Then, for more general MIMO channel, we consider the high SNR regime and compute the distortion SNR exponent. Guassian approximation of MIMO mutual information will also be used to derive an approximation for large the antenna system.

A. Single Antenna System (SISO)

For simplicity, we introduce the normalized delay as $\tau_n = \tau/T_f = \frac{1}{\theta B_w T_f} = \frac{1}{K\theta}$. For the SISO Rayleigh fading channel, the channel matrix degenerate to a scalar channel. We have following Corollary.

Corollary 1: For SISO system, the distortion-delay upper bound is

$$\mathcal{D}(\lambda\eta) \leq \left[\rho^{-\lambda\eta}\exp\left(\frac{1}{\rho}\right)\gamma\left(1-\lambda\eta,\frac{1}{\rho}\right)\right]^{\frac{1}{\lambda}},\tag{19}$$
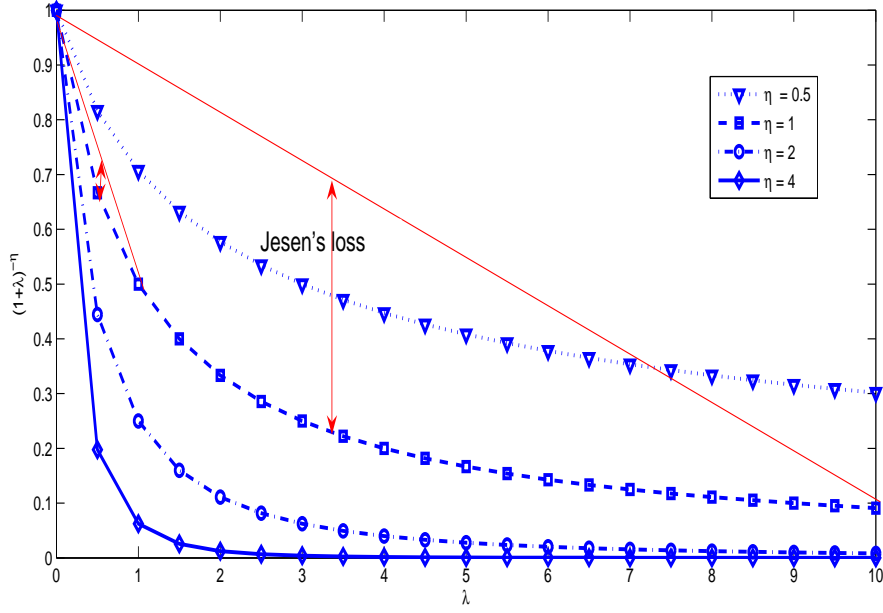
Fig. 3.   Illustration of buffer delay effect on the distortion

where $\lambda = \frac{1}{\tau_n}$ and $\gamma(\cdot, \cdot)$ is the incomplete Gamma function.

Proof: For SISO channel, the (14) is reduced to the scaler case,

$$\mathcal{D}(\lambda) \le \left[\int_0^\infty \left(1 + \rho x \right)^{-\lambda \eta} e^{-x} dx \right]^{\frac{l}{\lambda}} , \tag{20}$$

by the formula of [11], we can complete the proof.   ∎

The closed-form expressions of (19) is very difficult to analyze due to the special functions. In order to analyze distortion as the delay constraint increases, it is desirable to reduce the function into some simple form that is easy to handle. This objective motivates us to derive an asymptotically tight upper bound for the distortion-delay function in next section.

1) Asymptotic Analysis: We start by characterizing the behavior of the tail of distortion-delay curve $\mathcal{D}(\tau_n)$, hence we are interested in the asymptotically large delay regime. We will only consider Rayleigh fading SISO case. In this part, we assume $\eta = 1$ for simplicity, generalizing to other bandwidth ratio is straightforward. We try to show that $\mathcal{D}(\tau_n) \to \mathcal{D}(\infty)$ as $\tau_n \to \infty$. In addition, we will prove that the limit is approached as $e^{\frac{C}{\tau_n}}$ by finding the upper bound on the distortion-delay function and then show the bound is asymptotically tight. The ergodic capacity of

m_{th}-order diversity Raleigh fading channel with a constant transmission power can be expressed as [15]:

$$C_{erg} = \frac{\gamma(m, -m/\rho)}{\Gamma(m)} E_1(m/\rho) + \sum_{k=1}^{m-1} \frac{1}{k} \frac{\gamma(k, m/\rho)\gamma(m-k, -m/\rho)}{\Gamma(k)\Gamma(m-k)} , \qquad (21)$$

where $\gamma(\cdot, \cdot)$ and $\Gamma(\cdot)$ denote incomplete and complete Gamma functions; $E_1(\cdot)$ presents the exponential integration function. Hence for $m = 1$, the lower bound of distortion/delay function can be written as:

$$\mathcal{D}(\infty) = \exp\left(- e^{\frac{1}{\rho}} E_1(1/\rho)\right) . \qquad (22)$$

Next, We try to derive the asymptotic upper bound on $\mathcal{D}(\tau_n)$ of (19) to achieve the limit $\mathcal{D}(\infty)$. We mean asymptotically in the sense of $\tau_n \to \infty$ or $\lambda \to 0$.

Theorem 2: An asymptotic upper bound for $\mathcal{D}(D_n)$ can be expressed as:

$$\mathcal{D}_{upper}(\lambda) = \left[\frac{1}{\lambda - 1}\left(e^{\frac{1}{\rho}} - 1\right) + \frac{1}{1 - \xi\lambda + \phi\lambda^2}\rho^{-\lambda}e^{\frac{1}{\rho}} \right]^{\frac{1}{\lambda}} , \qquad (23)$$

where $\xi = 0.577215$ and $\phi = \frac{1}{12}(6\xi^2 - \pi^2)$. As $\lambda \to 0$ this upper bound is asymptotically tight and approaches $\mathcal{D}(\infty)$ as $\mathcal{D}(\infty) \cdot e^{C\lambda}$, where C is some constant.

Proof: See Appendix B. ∎

2) Example 1.: We present some numerical results to verify our findings. Suppose we have a real Gaussian source $N(0, 1)$ with bandwidth $100kHz$, bandwidth ratio $\eta = 1$[1]. We assume an i.i.d. block Rayleigh fading channel model. Let the duration of each time frame be $2ms$ such that each data frame consists of 200 source samples. Fig. 4 shows a normalized delay of $5T_f$ can achieve most of the gains, especially for high transmission power. The gap between this curve and the infinite delay case is less than 1dB for typical SNR value. In Fig. 5, we plot the end-to-end quadratic distortion vs. SNR and delay. It clearly characterizes the distortion and delay tradeoff for the Gaussian source transmitted over the wireless fading channel. Note that the higher the SNR value, the faster the distortion converges to the infinite delay lower bound. For SNR value of 25dB, less than $2T_f$ delay can achieve most of the Jensen's gain.

Fig. 6 shows the upper bound for the distortion/delay $\mathcal{D}(D_n)$ curve at SNR $= 15$dB. The ergodic Shannon capacity in this case is 3.0015 nats/symbol and the distortion $\mathcal{D}(\infty)$ is 0.0025. The rate of distortion/delay curve and the upper bound converge to the infinite delay lower bound

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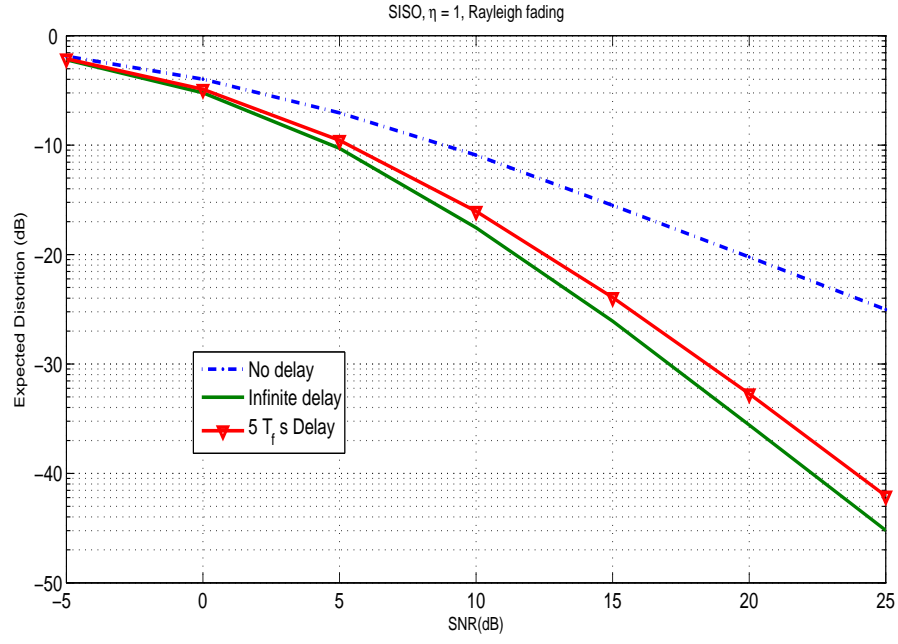[1]A real Gaussian source is equivalent to a complex one with doubled bandwidth ratio

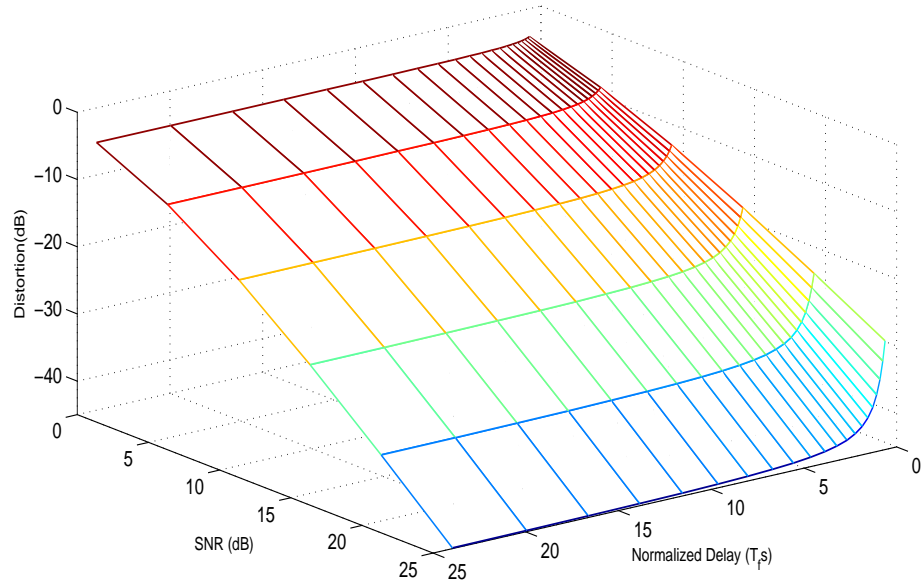Fig. 4. Distortion of Real Gaussian Source Transmitted over i.i.d. Rayleigh fading Channel.
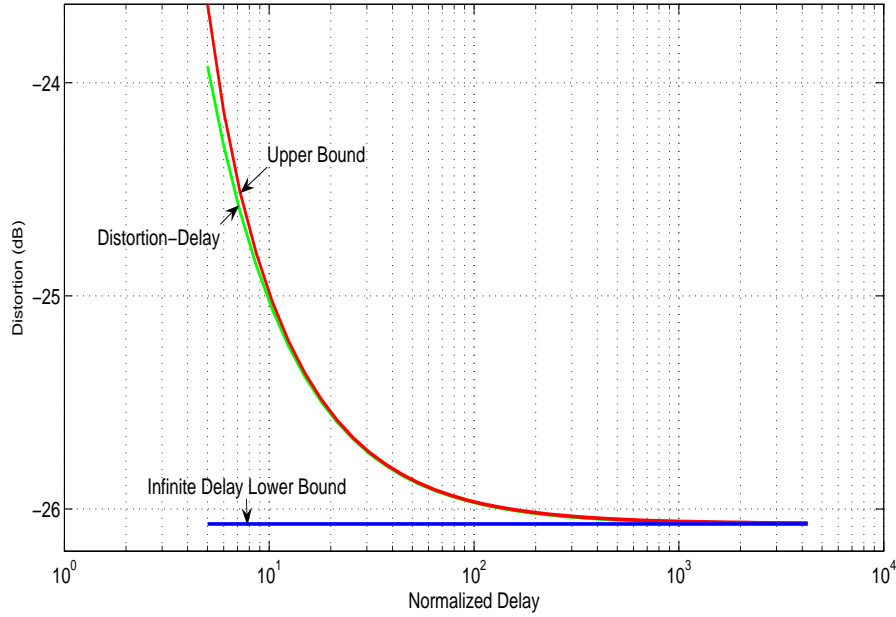

Fig. 5. Distortion vs. Delay and SNR

Fig. 6.   Upper Bound of Distortion/Delay function (SNR=15dB)

is clearly illustrated in Figure 5. It shows the upper bound is asymptotically tight and converges. From this upper bound and the distortion/delay function, we observe that introducing some finite delay can help achieving the $\mathcal{D}(\infty)$ lower bound very fast. In some practical applications, e.g., video transmission over wireless fading channel, which can tolerate certain amount of delay, our results suggest that inserting a buffer between quantizer and transmitter will enhance the image quality significantly. Intuitively, a transmission delay can be thought of as some delay diversity corresponding to space diversity in MIMO channel. Hence there is also some diversity-rate tradeoff for our problem, which can lead to results similar to those in [1].

B. SIMO/MISO Antennas System

For a SIMO channel of m receiver antenna. We can consider such channel as a m_{th}-order combining diversity Rayleigh fading channel. Again we here assume $\eta = 1$ for simplicity. The channel gain after combining is Chi-square distributed with $2m$ degrees of freedom, and the

probability density function (pdf) is given by:

$$f(x) = \frac{1}{(m-1)!} x^{m-1} e^{-x}, \quad x > 0 \ . \tag{24}$$

Corollary 2: For the SIMO Rayleigh fading channel with m receive antennas. The distortion-delay upper bound has a closed-form expression:

$$\mathcal{D}_m(\tau_n) \leq \left[\frac{\Gamma(\lambda - m)}{\Gamma(\lambda)} \rho^{-m} \, _1F_1\Big(m; m - \lambda + 1; \frac{1}{\rho}\Big) + \frac{\Gamma(m - \lambda)}{\Gamma(m)} \rho^{-\lambda} \, _1F_1\Big(\lambda; \lambda - m + 1; \frac{1}{\rho}\Big) \right]^{\tau_n}, \tag{25}$$

where $\lambda = 1/\tau_n$.

Proof: We start from Eqn. (14), with SIMO case

$$\begin{aligned}
\mathcal{D}(\theta) &= \Big(\int_0^\infty (1 + \rho x)^{-\lambda} f(x) d\rho \Big)^{\tau_n} \\
&= \Big(\frac{1}{(m-1)!} \cdot \int_0^\infty (1 + \rho x)^{-\lambda} x^{m-1} e^{-x} dx \Big)^{\tau_n},
\end{aligned} \tag{26}$$

where we have used the expression of $f(x)$ in (24). We know that [11, Ch. 3.383.5]:

$$\int_0^\infty e^{-px} x^{q-1} (1 + ax)^{-v} dx = a^{-q} \Gamma(q) \Psi\Big(q, q + 1 - v; \frac{p}{a}\Big) \ , \tag{27}$$

where $\Psi(\cdot, \cdot; \cdot)$ denotes the degenerate Hypergeometric function. Reducing to the more commonly used confluent hypergeometric function, we have following relation:

$$\Psi(x, y; z) = \frac{\Gamma(1 - y)}{\Gamma(x - y + z)} \, _1F_1(x; y; z) + \frac{\Gamma(y - 1)}{\Gamma(x)} z^{1-y} \, _1F_1(x - y + 1; 2 - y; z) \ . \tag{28}$$

Let $p = 1$, $q = m$, $v = \lambda$ and $a = \rho$. Plugging (28) into (27), we can prove Lemma 1.

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For MISO case[2], it is similar to the SIMO case but dividing the power by m. Even for the SIMO/MISO case the distortion-delay upper bound function is very complicate. We can only get some numerical results. Therefore, for more general MIMO channel, we resort to the SNR exponent in high SNR regime to demonstrate the buffer gain.

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[2]We assume transmitter has CSI for MISO case for beamforming transmission

V. Distortion Exponent of MIMO Block fading channel

For MIMO block fading channel with a buffered transmission, Eqn. (14) is very hard to analyze and provides less insight. We can only use the numerical method to compute the function. Since the "Jensen's gain" is negligible in low SNR regime and become significant at high SNR. Therefore we are more interested in the high SNR behavior of the expected distortion. We defined the figure of merit of *distortion exponent* [1] with bandwidth ratio η:

$$\alpha(\eta) = -\lim_{\rho \to \inf} \frac{\log \mathcal{D}(\rho, \eta)}{\log \rho} \ . \tag{29}$$

A distortion exponent of α means that the expected distortion decays as $\rho^{-\alpha}$ with increasing SNR value ρ when the SNR is high. We want to characterize the buffer delay and bandwidth ratio's impact on the SNR exponent.

Theorem 3: [1] (No Buffer) For transmission of memoryless, complex Gaussian source over a MIMO block fading channel, the distortion exponent with perfect known channel is given by

$$\alpha(\eta) = \sum_{i=1}^{M_*} \min \left(\eta, 2i - 1 + |M_t - M_r| \right). \tag{30}$$

The proof of Theorem 3, using the technique of [14]. Intuitively, when the bandwidth ratio is low, the distortion is limited by the η and the degree of freedom of MIMO channel – the total degree freedom utilized to transmit the information. One the other hand, when the bandwidth ratio is high, we need more diversity to provide the transmission reliability. Hence, for high bandwidth ratio, the system is diversity limited and the SNR exponent is determined by the second term.

Theorem 4: (with buffer delay) For transmission of memoryless, complex Gaussian source over a MIMO block fading channel, If the quantized bits are stored in a buffer before transmitting over the fading channel. Assume the transmitter know exactly the instantaneous channel capacity, the distortion SNR exponent is given by

$$\alpha(\eta) = \tau_n \min \left\{ \frac{\eta}{\tau_n}, 2i - 1 + |M_r - M_t| \right\} \ . \tag{31}$$

Proof: Proof can be found in Appendix II. ∎

Remarks

- We found the SNR exponent of Theorem 4 is similar as the one of joint encoding and decoding of L MIMO fading blocks. However, the joint encoding increase the transmitter
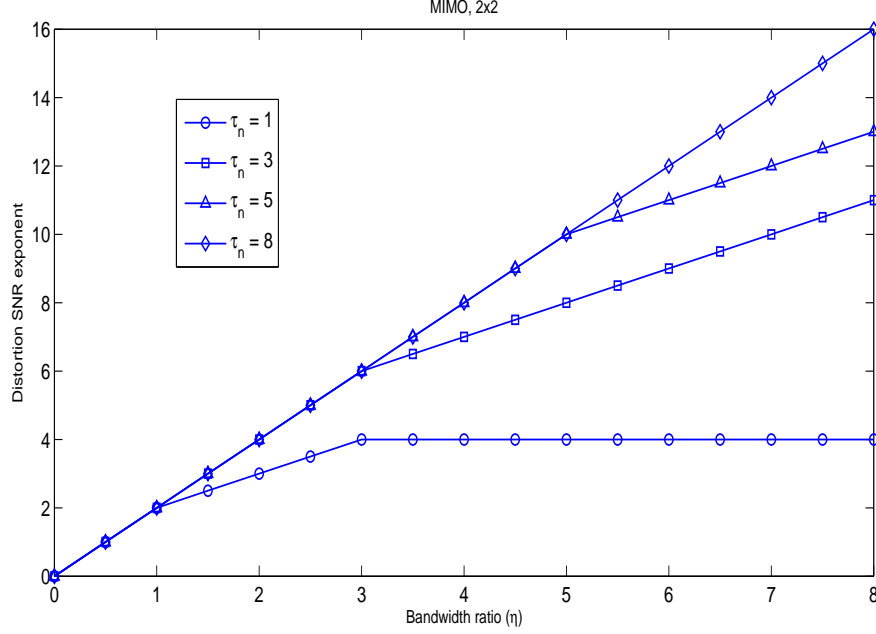
Fig. 7.   Distortion exponent v.s. bandwidth ratio for block fading 2x2 MIMO channel.

and receiver complexity. Introduce a simple buffer delay can get the same SNR exponent by utilizing the time diversity.

- For SIMO/MIMO case, the SNR exponent reduces to $\min\{\eta, \tau_n M\}$, where M is the receiver / transmitter antenna number. We can consider $\eta = \tau_n M$ as a corner point. Below this point, the system is degree of freedom limited, hence introduce more antenna will not improve the SNR exponent. Beyond this point, the system is diversity limited. Increasing the antenna number to provide more combining branches that will increase diversity, hence SNR exponent is also increased.

In Fig. 7, we fixed the MIMO channel as 2×2, and plotted the SNR exponent v.s. the bandwidth ratio curves for different delays. As the delay increases, we have more time diversity to combat fading, hence the corner point of the exponent-bandwidth ratio curve also increases. For $\tau_n = 1$, the maximum SNR exponent can be achieved for $\eta = 3$. It is useless to increase channel bandwidth ratio beyond 3 in the high SNR. In Fig. 8, We fixed the normalized delay as $\tau_n = 5$ and show different SNR exponent-bandwidth ratio curves for different antenna settings. For SISO channel, the SNR exponent will not increase anymore as the bandwidth ratio increase
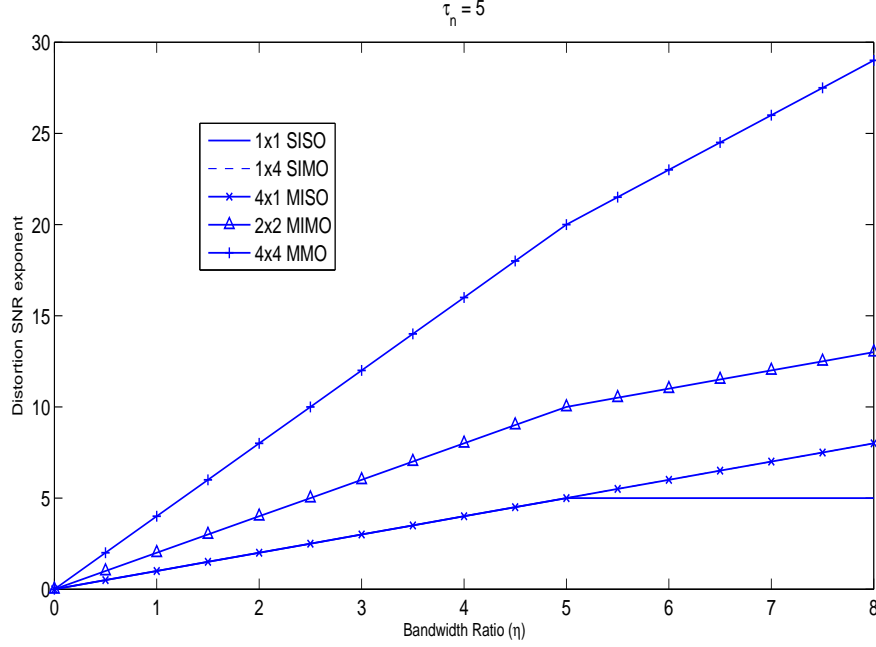
Fig. 8. Distortion exponent v.s. bandwidth ratio for normalized delay = 5.

beyond 5.

A. MIMO Mutual Information Gaussian Approximation

Due to inamenable to handle of Eqn. (14), we can use some approximations of the MIMO mutual information. The mathematical operation of $\log\det(\cdot)$ involves an extensive amount of average. Therefore the Lyapunov's central limit theorem can be applied. The mutual information can be approximate as a Gaussian distribution for large antenna systems. In [13], the mean and variance of different antenna settings has been derived. We will use the results of [13] to derive the distortion-delay approximations for different antenna settings.

1) Large M_r, fixed M_t: For this case the mutual information obeys

$$\mathcal{I} \sim \mathcal{N}\left(M_t\ln\left(1+\frac{M_r\rho}{M_t}\right), \frac{M_t}{M_r}\right). \tag{32}$$

The well-known moment generate function of the Gaussian distribution is $\mathrm{E}(e^{sx}) = \exp(sm_x + \frac{1}{2}s^2\sigma_x^2)$, where m_x and σ_x^2 is the mean and variance of the Gaussian variable x. Plug (32) into (9) and after some straightforward math manipulations, we can get the effective capacity and

distortion delay function as

$$E_c(\theta) = M_t \eta \ln\left(1 + \frac{M_r}{M_t}\rho\right) - \frac{1}{2}\theta K \frac{M_t}{M_r}\eta^2 \tag{33}$$

$$\mathcal{D}(\tau_n) \leq \left[1 + \frac{M_r\rho}{M_t} - \exp\left(\frac{M_t}{2M_r}(\frac{\eta^2}{\tau_n})\right)\right]^{-M_t\eta} \tag{34}$$

From Eqn. (33, 34), the effective capacity approaches to the ergodic capacity as $\theta \to 0$ or $M_r \to \infty$ (channel hardening). The SNR exponent is $M_t\eta$, which is the same as Theorem 4, as M_t fixed and M_r goes to infinity. Hence the SNR exponent is determined by the first term in Eqn. (31). We found the Guassian approximation did reveal the distortion-delay tradeoff asymptotically.

2) Large M_t, fixed M_r: the mutual information obeys

$$\mathcal{I} \sim \mathcal{N}\left(M_r \ln\left(1 + \rho\right), \frac{M_r\rho^2}{M_t(1+\rho)^2}\right). \tag{35}$$

The effective capacity and distortion delay curve is

$$E_c(\theta) = M_r \eta \ln(1+\rho) - \frac{1}{2}\theta K \eta^2 \frac{M_t}{M_r}\frac{\rho^2}{1+\rho^2} \tag{36}$$

$$\mathcal{D}(\tau_n) \leq \left[1 + \rho - \exp\left(\frac{M_r}{2M_t}(\frac{\eta^2}{\tau_n})\frac{\rho^2}{1+\rho^2}\right)\right]^{-M_r\eta} \tag{37}$$

Again, the effective capacity approaches to the ergodic capacity as $\theta \to 0$ or $M_t \to \infty$ The SNR exponent is $M_r\eta$, which confirmed the results of Theorem 4.

3) Large M_t and M_r, Fixed $\beta = M_r/M_t$, High SNR: The mutual information obeys

$$\mathcal{I} \sim \mathcal{N}\left(M_t\mu(\beta, \rho), \sigma^2(\beta)\right), \quad \beta \geq 1 \tag{38}$$

$$\sim \mathcal{N}\left(M_r\mu\left(\frac{1}{\beta}, \beta\rho\right), \sigma^2\left(\frac{1}{\beta}\right)\right), \quad \beta \leq 1. \tag{39}$$

Where $\mu(\beta, \rho) = \ln\rho + F(\beta)$, $F(\beta), \sigma^2(\beta)$ are functions only depends on β. The effective capacity capacity and distortion-delay function is:

$$E_c(\theta) = M_r \eta \ln(\rho) - \theta C_1 \tag{40}$$

$$\mathcal{D}(\tau_n) \leq \left[\rho - C_2\right]^{-M_r\eta}, \quad \beta \geq 1 \tag{41}$$

$$E_c(\theta) = M_t \eta \ln(\rho) - \theta C_3 \tag{42}$$

$$\mathcal{D}(\tau_n) \leq \left[\rho - C_4\right]^{-M_t\eta}, \quad \beta \geq 1, \tag{43}$$

Where C_1, C_2, C_3, C_4 are some constants. As both M_r, M_t goes to big and with fixed β, hence the $|M_t - M_r|$ also goes large, the SNR exponent is still $M_* \eta$.

VI. DISCUSSION AND REMARKS

In previous sections, we have clearly characterized the distortion/delay curve. However, we depend on some ideal assumptions, e.g., the instantaneous channel capacity is achievable and the CSI is perfectly known at the transmitter.

Remark 1: (*Decoding Error Probability*) In previous discussion we have assume using the the Gaussian code to achieve the instantaneous capacity. In reality, we have to take the decoding error probability into account for short codewords. [6] has integrated the physical layer decoding error into the effective capacity function through random coding error exponent. They have shown a joint queuing/coding exponent exits. Such an exponent can fit well into our distortion and delay analytical frame work.

Remark 2: (*Power Control*) Since we have perfect CSI at the transmitter, given an average transmission power budget, we can control the transmission power to maximize the effective capacity or minimize the end-to-end distortion for some delay constraint. In other words, the transmission power is not necessarily constant. Recent work [?] shows that, the optimum power adaptation policy is related to the delay constraint. As the delay goes to infinity, the power control policy approaches water-filling solution. On the contrary, for stringent delay constraints, the optimum power control policy becomes more like "truncated channel inversion". In the future work, we will investigate the how optimum power control affects the distortion/delay curves.

Remark 3: (*Channel Correlation*) Although i.i.d. block fading channel is easy to analyze and has several practical applications, this model is not always valid. It is more general and practical to consider channel correlation. We can use Jake's model to characterize the correlated channel fading process. The autocorrelation of channel gain $R(\tau)$ can be expressed as

$$R(\tau) = J_0(2\pi f_d \tau) \ , \tag{44}$$

where $J_0(\cdot)$ denotes the zero-th order Bessel function of first kind and f_d represents the maximum Doppler frequency. Channel correlation will reduce the effective capacity[5]. Intuitively, correlation may cause the fading channel to stay in the bad status for a longer time compared with i.i.d. block fading. [?] shows that given a correlated fading channel with the same marginal

statistics as i.i.d. case, the effective capacity of such a correlated channel is a linear shift in delay axis in logarithmic scale, the shift value is proportional to the Doppler frequency f_d. Hence the i.i.d. block fading distortion/delay tradeoff can be easily extended to the correlated case.

VII. CONCLUSION

We investigate the fundamental problem of distortion/delay tradeoff for the analogue source transmitted over wireless fading channels. We derive a close-form analytical formula to characterize this relationship using recently proposed effective capacity. Based on this closed-form expression, we give out an upper bound that is asymptotically tight to study the convergence behavior of the distortion/delay function for SISO channel. We also characterized the SNR exponent of MIMO block fading channel in the high SNR regime. Simulation results show that a small delay can result in a significant transmission power save. The framework of this paper is applicable to a broad class application, e.g., video transmission.

APENDIX A. PROOF OF THEOREM 2

Proof: From Eqn. (19) of Corollary, we have

$$
\mathcal{D}(\lambda) \leq \left[\rho^{-\lambda} \exp\left(\frac{1}{\rho}\right) \gamma\left(1 - \lambda, \frac{1}{\rho}\right) \right]^{\frac{1}{\lambda}}
$$

$$
= \left[\frac{1}{\lambda - 1} \frac{1}{\rho} \,_1F_1\left(1; 2 - \lambda; \frac{1}{\rho}\right) + \Gamma(1 - \lambda)\left(\frac{1}{\rho}\right)^{\lambda} \exp\left(\frac{1}{\rho}\right) \right]^{\frac{1}{\lambda}} \tag{A-1}
$$

Since $\frac{1}{\lambda-1} < 0$ as $\lambda \to 0$, we first lower-bound the confluent hypergeometric function.

$$
_1F_1(1; 2 - \lambda; x) = \sum_{k=0}^{\infty} \frac{(1)_k}{(2 - \lambda)_k} \frac{x^k}{k!}
$$

$$
\geq \sum_{k=0}^{\infty} \frac{(1)_k}{(2)_k} \frac{x^k}{k!} = \frac{1}{x}(e^x - 1) , \tag{A-2}
$$

where $(a)_k \triangleq a \cdot (a + 1) \cdots (a + k - 1)$. For $\lambda \to 0$ this lower bound is asymptotically tight. Next we upper-bound the $\Gamma(1 - \lambda)$.

$$
\Gamma(1 - \lambda) = -\lambda \cdot \Gamma(-\lambda) = \frac{-\lambda}{\frac{1}{\Gamma(-\lambda)}}
$$

$$
= \frac{-\lambda}{-\lambda + \xi(-\lambda)^2 + \phi(-\lambda)^3 + \delta(-\lambda)^4 + O((-\lambda)^5)}
$$

$$
\leq \frac{1}{1 - \xi\lambda + \phi\lambda^2 - \delta\lambda^3} , \tag{A-3}
$$

where $\xi = 0.577215$, $\phi = \frac{1}{12}(6\xi^2 - \pi^2)$ and δ is some constant. Hence replacing (A-2) and (A-3) in (??) we have the following upper bound

$$\mathcal{D}(\lambda) \dot{\leq} \left[\frac{1}{\lambda - 1} \left(e^{\frac{1}{\rho}} - 1 \right) + \frac{1}{1 - \xi\lambda + \phi\lambda^2} \rho^{-\lambda} e^{\frac{1}{\rho}} \right]^{\frac{1}{\lambda}} , \tag{A-4}$$

where we have omitted $O(\lambda^3)$ term, which will not affect the result as $\lambda \to 0$. Using Taylor expansion for the first term and second term, and dropping the $O(\lambda^3)$, we obtain the following asymptotic approximation,

$$\mathcal{D}_{upper}(\lambda) \dot{\approx} [1 + a\lambda + b\lambda^2]^{\frac{1}{\lambda}}$$
$$= \exp(a) \exp\left((b - \frac{a^2}{2})\lambda \right) , \tag{A-5}$$

where we have used the identity $\lim_{x \to 0}(1 + x)^{\frac{1}{x}} = e$, and

$$a \triangleq 1 - e^{\frac{1}{\rho}} + \xi e^{\frac{1}{\rho}} - \ln \rho e^{\frac{1}{\rho}}$$
$$b \triangleq 1 - e^{\frac{1}{\rho}} + (\xi^2 - \phi)e^{\frac{1}{\rho}} - \xi \ln \rho e^{\frac{1}{\rho}} + \ln^2 \rho .$$

In order to show $\mathcal{D}_{upper}(\lambda) \to \mathcal{D}(\infty)$ in (22), in other word (A-5) $\to$ (22), we want to show that

$$F \triangleq 1 - e^{-\frac{1}{\rho}} - \xi + \ln \rho \to E_1(1/\rho) . \tag{A-6}$$

$E_1(\cdot)$ is a special function, and don't have simple expression. Instead we use numerical method to illustrate the convergence. We have plotted these two values in Figure 6. We can observe for most SNR these two values match perfectly. Hence we conclude that the upper bound converges and the convergent rate is exponential. ∎

APPENDIX B. PROOF OF THEOREM 4

Proof: We will follow the technique used in [14]. Assume without loss of generality that $M_t = M_* \leq M_r$ (the case $M_t > M - r$ is a simple extension). We start from the distortion delay function (17)

$$\mathcal{D}(\rho) = \left\{ \int_0^\infty \prod \left(1 + \frac{\rho}{M_t}\lambda_i \right)^{-\theta K\eta} f(\boldsymbol{\lambda})d\boldsymbol{\lambda} \right\}^{\frac{1}{\theta K}} , \tag{A-7}$$
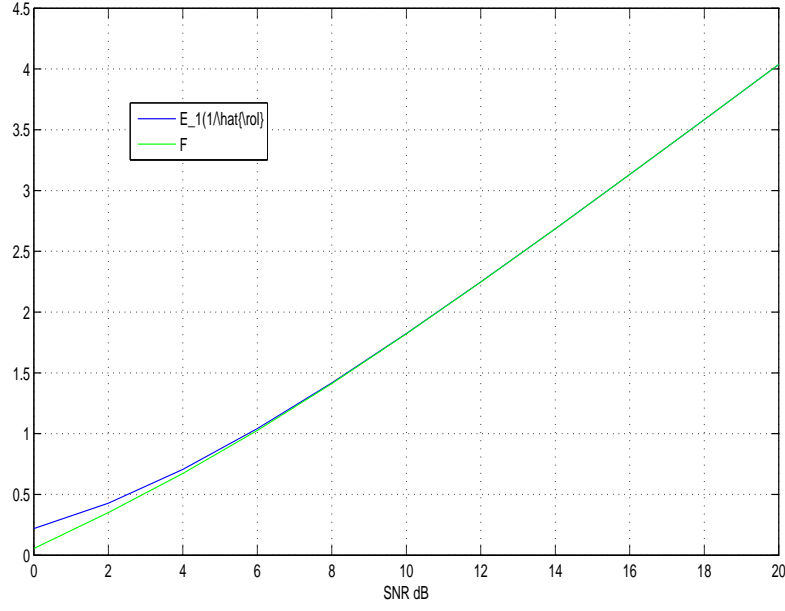
Fig. 9.   Illustration (A-8) for different SNR values

where $\lambda_1 \leq \lambda_2 \leq \cdots \leq \lambda_{M_t}$ are the ordered eigenvalues of $\mathbf{H}\mathbf{H}^H$. We make the change of variable: $\alpha_i = -\log(\lambda_i)/\log(\rho)$, for all $i = 1, \cdots, M_t$, The joint pdf $\boldsymbol{\alpha} = [\alpha_1, \cdots, \alpha_{M_t}]$, where $\alpha_1 \geq \cdots \geq \alpha_{M_t}$, is given by

$$f(\boldsymbol{\alpha}) = K_{M_t,M_r}^{-1} \left(\log \rho \right)^{M_t} \prod_{i=1}^{M_t} \rho^{-(M_r - M_t + 1)\alpha_i} \prod_{i<j} \left(\rho^{-\alpha_i} - \rho^{-\alpha_j} \right)^2 \exp \left(\sum_i \rho^{-\alpha_i} \right) . \qquad \text{(A-8)}$$

Replace $\boldsymbol{\lambda}$ with $\boldsymbol{\alpha}$, (A-7) yields

$$\mathcal{D}(\rho) = \left\{ \int_{\mathcal{A}} \prod_{i=1}^{M_t} (1 + \frac{1}{M_t} \rho^{1-\alpha_i})^{-\theta K \eta} f(\boldsymbol{\alpha}) d\boldsymbol{\alpha} \right\}^{\frac{1}{\theta K}} , \qquad \text{(A-9)}$$

where

$$\mathcal{A} = \left\{ \boldsymbol{\alpha} \in \mathbb{R}^{M_t} \quad : \quad \alpha_1 \geq \cdots \geq \alpha_{M_t} \right\} .$$

Neglecting all terms that irrelevant to the SNR exponent, we obtain (A-7) yields

$$
\begin{aligned}
\mathcal{D}(\rho) &\dot{\geq} \left\{ \int_{\mathcal{A} \bigcap \mathbb{R}^{M_t}+} \left(\prod_{i=1}^{M_t}(1 + \frac{1}{M_t}\rho^{1-\alpha_i})^{-\theta K\eta} \right) \prod_{i=1}^{M_t} \rho^{-(2i-1+M_r-M_t)\alpha_i} d\boldsymbol{\alpha} \right\}^{\frac{1}{\theta K}} \\
&\dot{=} \left\{ \int_{\mathcal{A} \bigcap \mathbb{R}^{M_t}+} \prod_{i=1}^{M_t} \rho^{-\theta K\eta(1-\alpha_i)^+} \prod_{i=1}^{M_t} \rho^{-(2i-1+M_r-M_t)\alpha_i} d\boldsymbol{\alpha} \right\}^{\frac{1}{\theta K}} \\
&\dot{=} \left\{ \int_{\mathcal{A} \bigcap \mathbb{R}^{M_t}+} \prod_{i=1}^{M_t} \rho^{-(\theta K\eta(1-\alpha_i)^+ + (2i-1+M_r-M_t)\alpha_i)} \right\}^{\frac{1}{\theta K}} \\
&\dot{=} \rho^{\alpha(\eta)\frac{1}{\theta K}}
\end{aligned}
\tag{A-10}
$$

where we have used

$$
(1 + \frac{1}{M_t}\rho^{1-\alpha_i})^{-\theta K\eta} \dot{=} \rho^{-\theta K\eta[1-\alpha_i]^+} .
$$

And

$$
\alpha(\eta) = \inf_{\boldsymbol{\alpha} \in \mathcal{A} \bigcap \mathbb{R}^{M_t}+} \sum_{i=1}^{M_r} (2i - 1 + M_r - M - t)\alpha_i + \theta K\eta(1 - \alpha_i)^+ .
$$

We can minimizing individual term of the summation separately by set $\alpha_i = 0$ or 1. We also notice that $\theta K = \tau_n$, the buffer delay, hence we can obtain the SNR exponent of the buffered transmission is

$$
\alpha(\eta) = \tau_n \min \left\{ \frac{\eta}{\tau_n}, 2i - 1 + M_r - M_t \right\} .
\tag{A-11}
$$

■

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